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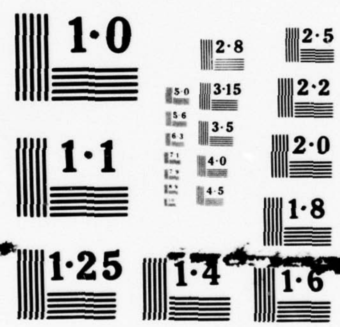
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LOGARITHMIC METHOD OF GENERALIZED AMPLITUDE DETECTION

by

V. N. Nogin

FOREIGN TECHNOLOGY DIVISION

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Block	Italic	Transliteration	Block	Italic	Transliteration
А а	*А а*	A, a	Р р	*Р р*	R, r
Б б	*Б б*	B, b	С с	*С с*	S, s
В в	*В в*	V, v	Т т	*Т т*	T, t
Г г	*Г г*	G, g	У у	*У у*	U, u
Д д	*Д д*	D, d	Ф ф	*Ф ф*	F, f
Е е	*Е е*	Ye, ye; E, e*	Х х	*Х х*	Kh, kh
Ж ж	*Ж ж*	Zh, zh	Ц ц	*Ц ц*	Ts, ts
З з	*З з*	Z, z	Ч ч	*Ч ч*	Ch, ch
И и	*И и*	I, i	Ш ш	*Ш ш*	Sh, sh
Й й	*Й й*	Y, y	Щ щ	*Щ щ*	Shch, shch
К к	*К к*	K, k	Ъ ъ	*Ъ ъ*	"
Л л	*Л л*	L, l	Ы ы	*Ы ы*	Y, y
М м	*М м*	M, m	Ь ь	*Ь ь*	'
Н н	*Н н*	N, n	Э э	*Э э*	E, e
О о	*О о*	O, o	Ю ю	*Ю ю*	Yu, yu
П п	*П п*	P, p	Я я	*Я я*	Ya, ya

*<u>ye</u> initially, after vowels, and after ъ, ь; <u>e</u> elsewhere.
When written as ё in Russian, transliterate as yё or ё.

RUSSIAN AND ENGLISH TRIGONOMETRIC FUNCTIONS

Russian	English	Russian	English	Russian	English
sin	sin	sh	sinh	arc sh	$\sinh^{-1}$
cos	cos	ch	cosh	arc ch	$\cosh^{-1}$
tg	tan	th	tanh	arc th	$\tanh^{-1}$
ctg	cot	cth	coth	arc cth	$\coth^{-1}$
sec	sec	sch	sech	arc sch	sech^{-1}
cosec	csc	csch	csch	arc csch	csch^{-1}

Russian	English
rot	curl
lg	log

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LOGARITHMIC METHOD OF GENERALIZED AMPLITUDE DETECTION.

V. N. Nogin, graduate student.

The problem of closeness in "ether/ester" and the freedom from interference of radio reception advanced into turn the task of the separation of radio signals with the overlapping spectra. During its solution sometimes appears the need for detection of such amplitude-modulated (AM) oscillation/vibrations, "carrier" frequency of which can have any value up to zero hertzes. In fact, let, for example, we have a sum two AM of the signals

$$u_\Sigma = U_1(t)\sin(\omega_1 t + \psi_1) + U_2(t)\sin(\omega_2 t + \psi_2).$$

On Fig. 1a are depicted their spectra. After synchronous (or selective [1]) quadrature detection at carrier frequency ω_2 jamming station, i.e., the multiplication of the total oscillation/vibration

by the voltage of auxiliary heterodyne $U_r \cos(\omega_2 t + \psi_2)$ and the filtrations of all high-frequency component we will obtain

$$u_1 = U(t) \sin(\Omega t + \Delta\psi). \quad (1)$$

Formally expression (1) is AM oscillation of the adopted station, but with the lowered/reduced "carrier" frequency Ω. In it $U(t) = \frac{1}{2} U_r \cdot U_1(t)$ — "the is envelope" of the carrier amplitudes oscillation; $\Omega = \omega_1 - \omega_2$ and $\Delta\psi = \psi_1 - \psi_2$ - with respect to a difference in frequencies and initial carrier frequencies the component initial AM of the signals. If the spectra of the latter overlap (Fig. 1a), then frequency Ω can lie/rest at the range of audio frequencies and even lower. In common AM oscillations the carrier frequency considerably higher than highest frequency of the modulating. AM of the oscillations, in which this condition is not observed, i.e., on carrier frequency are placed no limitations, let us call those generalized by AM oscillations.

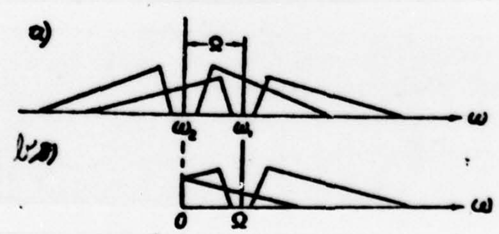

Fig. 1.

Fig. 1. Spectra: a) two AM of the signals; b) generalized AM of oscillation.

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On Fig. 1b is depicted the amplitude spectrum of this oscillation obtained by the way of the described synchronous detection from spectrum Fig. 1a. Its lower side band as is bent relative to zero frequency. The real spectrum of the bent part stretches to the side of positive frequencies [2; 3], partially overlapping with the spectrum of upper side band. To Fig. 2 is shown the cut variable oscillogram of this generalized AM of oscillation. As we see, it can be and not is similar to oscillation with the slowly changing carrier amplitude. Therefore the common methods of detection, based on the isolation/liberation of the modulating signal of as envelope of the carrier amplitudes oscillation, are here unsuitable.

From expression (1) it is directly evident that for the generalized amplitude detection, i.e., theisolation/liberation "of the envelope" generalized AM of oscillation, the latter is sufficient to divide by the harmonic oscillation, synchronous and cophasal with "carrying" variation of frequency Ω. This can be carried out one of the methods of functional division, for example, described in [4-7].

The harmonic voltage, cophasal with "carrier", can be separated either by the narrow-band filter from spectrum Fig. 1b, tuned to a frequency Ω or to shape by filterless method from the oscillation of rectangular form (Fig. 2b), the obtained by means amplification - the limitation of generalized AM of oscillation (Fig. 2a). The latter is possible because this rectangular oscillation includes the information about the phase of "carrying" oscillation, since the torque/moments of the transition through zero in them coincide.

Let us examine now the logarithmic method of detection the generalized of AM oscillations, which virtually render/showed simpler methods of division indicated. Let us record expression (1) in the form

$$u_1 = U_0[1 + mf(t)]\sin(\Omega t + \Delta\psi), \qquad (2)$$

where f (t) - the quasi-periodic function, which is the standardized/normalized modulating oscillation, i.e., the sound vibration, referred to the average modulus of its instantaneous values; m — the average coefficient of the depth of modulation.

Let us manufacture full-wave rectification for active load and the subsequent [1] logarithmic operation of generalized AM of the

oscillation (Fig. 2a), presented by expression (2).

FOOTNOTE 1. Of the operations of rectification and logarithmic operation it is possible and to interchange the position, but then logarithmic amplifier must have symmetrical amplitude characteristic. ENDFOOTNOTE.

We will obtain

$$u_{1\,\text{лог}} = \log_N \{U_0 [1 + mf(t)] \cdot |\sin(\Omega t + \Delta\psi)|\} =$$
$$= \log_N U_0 + \log_N |\sin(\Omega t + \Delta\psi)| + \log_N [1 + mf(t)]. \quad (3)$$

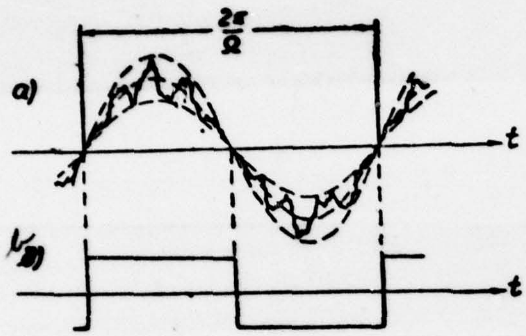

Fig. 2. Oscillograms: a) generalized AM of oscillation; b) rectangular oscillation.

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The process of ideal logarithmic operation is illustrated on Fig. 3a. In expression (3) the first term is constant component and therefore us it does not interest. The second term represents the periodic oscillations u_Δ (Fig. 3a), which is present at the output/yield of the logarithmator as the additive mixing oscillation. Its expansion in Fourier series takes form [8]:

$$u_\Delta = \log_N |\sin(\Omega t + \Delta\psi)| = -a\left[\ln 2 + \sum_{k=1}^{\infty} \frac{1}{k} \cos(2k\Omega t + 2k\Delta\psi)\right],$$

i.e. it consists of the even harmonics of frequency Ω (Fig. 3b). Here $a = 1/\ln N$ is a module/modulus of transition to natural logarithm.

The third term in expression (3) it is possible to decompose in the following power series, which is converged with $|mf(t)| < 1$, which virtually always is made.

$$u'_{1\,\text{лог}} = a \ln [1 + m f(t)] =$$
$$= a \left[m f(t) - \frac{m^2}{2} f^2(t) + \frac{m^3}{3} f^3(t) - \frac{m^4}{4} f^4(t) + \dots \right]. \qquad (4)$$

Here the first term in brackets represents the pure/clean sound vibration, obtained as a result of the logarithmic method of detection. Its spectrum S (ω) is depicted on Fig. 3b. The terms of higher degrees testify to the presence of the undesirable components, which are the products of nonlinear distortions during detection, which appear as a result of the nonlinearity of logarithmic characteristic.

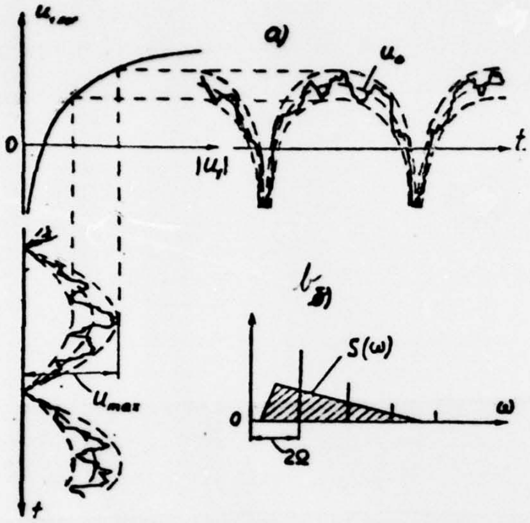

Fig. 3. **Logarithmic operation: a) the ideal logar**ithmic operation of rectified generalized AM of oscillation; b) oscillation spectrum at the output/yield of the logarithmator.

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From expression (4) it is evident that with low m the terms of higher degrees can be disregarded and then Fig. 3b with high accuracy/precision it represents the spectrum of full wave at the output/yield of the logarithmator. Because of logarithmic operation the product is converted into the sum, from which it is possible to separate the modulating signal, for which the additive mixing

oscillation u_Δ can be either filtered out by the blocking comb filter, which does not pass frequencies, multiple 2Ω or it is compensated for. In the latter case for obtaining the compensating oscillation it is necessary to separate the "carrying" oscillation, to rectify it on resistive load and to take the logarithm. The comb blocking filter can be carried out on the base of the circuits of cross-period compensation [9]. Apparently, the method of this filtration is unsuitable at the values of the frequencies Ω, for which the half fundamental frequency of the sound (modulating) signal is multiple, since otherwise could be filtered out all the spectral components of useful signal. However, with voice signals these values Ω do not exceed the limits of a comparatively narrow band of frequencies 40-150 Hz (Ω/2π).

Let us pass to the quantitative estimate/evaluation of the nonlinear distortions, which appear as a result of nonlinearity logarithmic characteristic. For the purpose of simplification let us define, as this is usually accepted, the coefficient of harmonics. Let f (t) is be the sinusoidal function of single amplitude. Then sum n of the first terms of series (4) will be the exponential trigonometric polynomial, converting [8] which in harmonic trigonometric polynomial can be found K_r. Being limited to the number of terms n = 4 and by lowering intermediate computations, let us give the resultant expression for the total coefficient of the second and

third harmonics

$$K_{\rm r} \approx \frac{m\sqrt{9(2+m^2)^2+4m^2}}{6(4+m^2)}. \tag{5}$$

The coefficient of harmonic nonlinear distortions during detection (Fig. 4, curve 1) are almost directly proportional to the depth of modulation, whereupon it is comparatively small and with the standard depth of modulation m = 30o/o does not exceed 8o/o.

It is noted [10] that the amplitude characteristic of real logarithmator differs from that depicted by Fig. 3a as presence of the initial linear section whose value on input voltage let us designate by $u_{\rm n}$.

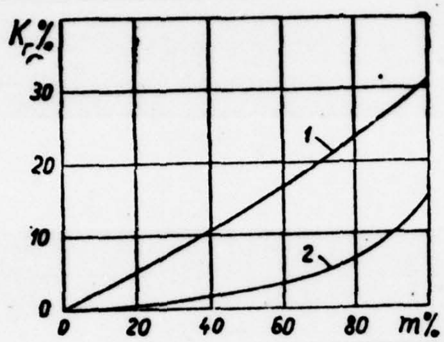

Fig. 4. Dependence of the coefficient of harmonics, caused by the nonlinearity of logarithmic characteristic, on the depth of modulation: 1 - for the asymmetric version; 2 - for a symmetrical version.

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Due to this sonic output potential of this detector had been have as if narrowings at the points of the passage through zero of detected oscillation, i.e., sound vibration will have certain envelope. The formation of the latter it explains Fig. 5. On the upper part of the figure on the amalgamated scale is depicted the range of the initial values of voltage on the input of the logarithmator, in which the initial cuts of sinusoids (inclined dotted line) are close to straight lines. The amplitudes of sound vibration at output/yield for

the sections, where input voltage is **less** $u_н$, **grow/rise** linearly (sections oa and oг). On those sections (**аб and** гд), where the corresponding half-periods of sound vibration in the input signal intersect boundary/interface of MN between the linear and logarithmic sections of the characteristics, enveloping change nonlinear. From Fig. 5 it is evident that the envelopes on top (oaбв) and from below (oгде) are obtained differently. However, since the distortions in the ranges of the flat/plane sections of envelopes (бв and де), that appear as a result of the nonlinearity of logarithmic characteristic, already were taken into account above, with sufficient accuracy/precision to evaluate distortions due to narrowings envelopes from below and can be approximated on top by one and the same equilateral trapezium with the duration of inclined sides in time Δ/2.

Let us approach toward the quantitative estimate/evaluation of the distortions, caused by narrowings. Thus, we consider that the envelope of the sound signal is the periodic sequence of the trapezoidal momentum/impulse/pulses (Fig. 6) of single amplitude. Spectral representation for this function [8] can be recorded in the following form:

$$F(t) = 1 - \frac{\xi}{2} - \frac{4}{\pi^2}\sum_{n=1}^{\infty}\frac{\sin^2 n\frac{\pi}{2}\xi}{n^2\xi}\cos n\Omega_i t. \qquad (6)$$

Here $\xi = \frac{\Delta}{T_i}$ — the relative duration of narrowings; $\Omega_i = \frac{2\pi}{T_i}$ — the

frequency of their sequence.

Let the sound signal is be the complex periodic function

$$u_{зв} = \sum_{k=1}^{m} B_k \cos(k\Omega_{зв} t + \varphi_k). \tag{7}$$

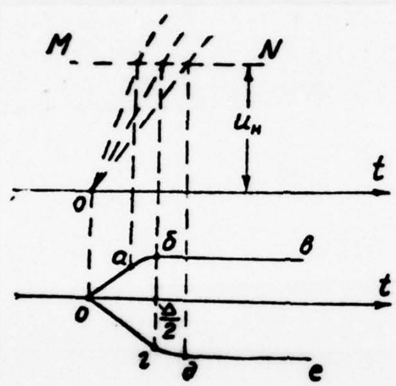

Fig. 5. Formation of narrowings in the output signal.

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Then taking into account (6) this same signal, subjected to narrowings, is recorded:

$$u_{вых} = u_{зв} \cdot F(t) = \left(1 - \frac{\xi}{2}\right) \sum_{k=1}^{m} B_k \cos(k\Omega_{зв} t + \varphi_k) -$$
$$- \frac{2}{\pi^2 \xi} \sum_{n=1}^{\infty} \sum_{k=1}^{m} \frac{\sin^2 n\frac{\pi}{2}\xi}{n^2} B_k \{\cos[(n\Omega_i + k\Omega_{зв})t + \varphi_k] +$$
$$+ \cos[(n\Omega_i - k\Omega_{зв})t - \varphi_k]\}. \tag{8}$$

Comparing the last/latter expression with (7), it is not difficult to note that the first term here represents the pure/clean sound signal. Its amplitude because of narrowings somewhat decreased. Relative decrease it composes $\varepsilon/2$. The second and third terms of expression (8) are the spectral components of combination frequencies, which arose due to the narrowings of the signal. These are the products of nonlinear distortions.

Since sonic $u_{зв}$ and enveloping F (t) of oscillation are independent, $\Omega_{зв}$ and Ω_i in the majority of cases they will not be found in affine multiple ratios. Consequently, it is possible to count that the different components of expression (8) do not coincide in frequency and therefore occurs the add of the powers of separate spectral components.

The coefficient of nonlinear distortions will be greatest, when only the negligible part of the combination frequencies in expression (8) exceeds the limits of the spectrum of the modulating oscillation and therefore can be filtered out. This will be at comparatively low repetition frequencies of narrowings Ω_i, i.e., with $\Omega_i \ll \Omega_{зв\,max}$. For this heaviest case let us determine on the basis (8) the coefficient of nonlinear distortions because of the narrowings:

$$\gamma_{D\,max} = \frac{\sqrt{2\sum_{n=1}^{\infty}\left(\frac{2}{\pi^2\xi}\cdot\frac{\sin^2 n\frac{\pi}{2}\xi}{n^2}\right)^2}}{1-\frac{\xi}{2}}. \qquad (9)$$

From (9) we see that $\gamma_{D\max}$ **does not depend on the form of modulating (sonic) signal (7). This essentially differs the nonlinear** distortions of this form from those which appear because of the nonlinearity of amplitude characteristic.

Is simplified the recording of expression (9)

$$\gamma_{D\max} = \frac{4\sqrt{2}}{\pi^2 \xi (2-\xi)}\sqrt{\sum_{n=1}^{\infty}\frac{\sin^4 n\frac{\pi}{2}\xi}{n^4}}.$$

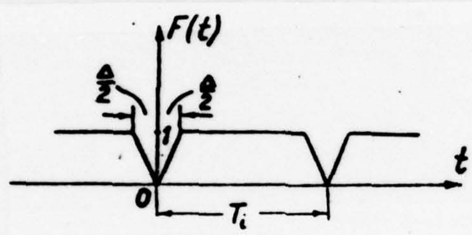

Fig. 6.

Approximation of the input signal envelope by a equilaternal trarezium.

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After expression under the root of the degree of the sine through the functions of the multiple arguments infinite sum decays to three, for each of which in [11] are expressions in the convolute form. Finally **we will obtain**

$$\gamma_{D\max} = \frac{1}{2-\xi}\sqrt{\xi\left(\frac{4}{3}-\xi\right)}. \qquad (10)$$

However, the relative duration of narrowings ξ **is unambiguously connected with the dynamic range D [10] of the logarithmic section of** the usable portion of the amplitude characteristic of the **logarithmator. Actually, on the basis Fig. 3a and Fig. 5 it is possible to write**

$$D = \frac{u_{max}}{u_n} = \frac{1+m}{\sin \Omega \frac{\Delta}{2}}.$$

After considering that $\Omega = \frac{1}{2}\Omega_i$, and after requiring in order that the input voltage of the logarithmator with any m would not exceed the upper limit of the logarithmic section of characteristic, i.e., set/assuming $m = m_{max} = 1$, from the last/latter expression let us find

$$\xi = \frac{2}{\pi} \arc \sin \frac{2}{D}. \quad (11)$$

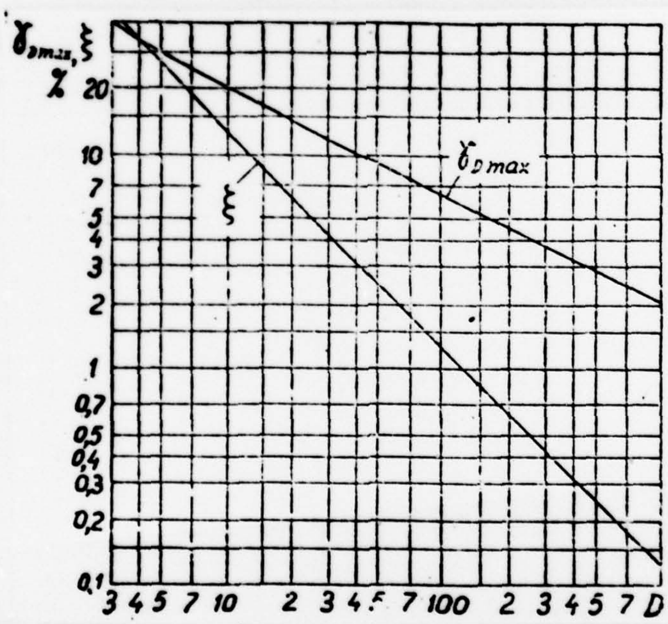

Fig. 7. Dependence of the relative duration of narrowings ξ and of the caused by them maximum coefficient of combination distortions $\gamma_{D\,max}$ from the dynamic range of loarithmator D.

By formulas (10) and (11) are designed the graph/diagrams of the dependences $\gamma_{D\ max}$ and ξ on D (Fig. 7). Let us note that in actuality m rarely it reaches one. Therefore values D, obtained from graph Fig. 7, one should consider as maximum. Virtually it is possible to take somewhat smaller values of D.

To evaluate the total nonlinear distortions due to the constrictions and due to the nonlinearity of the logarithmic units amplitude characteristic they should be expressed uniformly. Although the products of distortions in expression (8) also does not contain the signal harmonic, in order to employ the widely propagated estimate of nonlinear distortions in accordance with the coefficient of harmonics, we arbitrarily replace distortions due to constrictions by equivalent (for γ) distortions due to the nonlinearity of amplitude characteristic. In this case let us consider that the coefficient of the combination distortions of the sound signal, which appear as a result of the nonlinearity of amplitude characteristic, several (ν) once (ν depends on waveform) more the coefficient of harmonics [12]. It means the equivalent coefficient of harmonics, which conditionally characterizes distortions due to narrowings, it will be several times less than $\gamma_{D\ max}$. But since the distortions of two forms in question are not depended mutually, it occurs the summation of the powers of distortion products. Then the total coefficient of the harmonics

$$K'_{\Gamma\,\text{общ}} = \sqrt{K_{\Gamma}^2 + \left(\frac{\gamma_{D\,max}}{\nu}\right)^2}. \tag{12}$$

Hence it is apparent that if we assume $\gamma_{D\,max}$ equal to K_{Γ}, then $K'_{\Gamma\,\text{общ}}$ insignificantly will exceed value K_{Γ}. So with $\nu = 2$ Means it is possib le to allow $\gamma_{D\,max}$ equal to the average value K_{Γ}, i.e., 8%. Then from graphs Fig. 7 we find that required D do not exceed 60, but the relative duration of narrowings ξ in this case will be about 2.1%. It is noted [10] that the manufacture of logarithmators with this dynamic range, which does not exceed two orders, represents no difficulties.

Let us explain now requirements for the accuracy/precision of the amplitude characteristic of the logarithmatcr. In its our case it is convenient to rate/estimate not by relative error δ [10] logarithmic characteristic, but by relative error ε the characteristic of the differential gear ratio of the logarithmator. since this value directly characterizes the degree of signal distortion due to an inaccuracy in the logarithmic characteristic. Furthermore, error ε can be measured incomparably faster and more precise than δ, since for its determination it is not required to take the graph of amplitude characteristic. These two errors of ideal logarithmator $(D \to \infty)$ are equal, and of real $\delta < \varepsilon$. The differential gear ratio of the logarithmator, as is known, directly proportional to the module/modulus of transition a (see above), that characterizes the slope/inclination of logarithmic amplitude characteristic. Therefore value ε can be also considered as a relative error in coefficient a.

Thus, let the characteristic of the differential gear ratio of the logarithmator has relative deflections from the ideal within limits $\pm\varepsilon$. It is clear that because of this the amplitude of useful sonic output potential of the logarithmator (see formula (4) ~~A~~ and Fig 3a) will also have relative deflections from its normal value on $\pm\varepsilon$.

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In other words, output voltage will turn out to be that modulated in amplitude with the depth of modulation, equal to ε. This modulation - idle, and its law is unknown. However, it will be its knowingly periodic; therefore standardized/normalized envelope (referred to the average value) it can be decomposed in Fourier series with the amplitude factors

$$A(k)=1+\sum_{k=1}^{\infty}A_k. \qquad (13)$$

By comparing (13) and (6), by analogy with (9) let us record expression for the coefficient of nonlinear distortions because of an inaccuracy in the amplitude characteristic of the logarithmator

$$\gamma_\varepsilon=\sqrt{2\sum_{k=1}^{\infty}\left(\frac{A_k}{2}\right)^2}.$$

From the last/latter expression we see that γ_ε is equal to the RMS value of the variable component of the standardized/normalized enveloping oscillation. Its amplitude is known and equal to ε. It is known that of all possible forms of the oscillations, which have the assigned amplitude ε, the RMS value maximally is equal to amplitude of square oscillations. However, the possibility of this envelope shape is in practice excluded, vsince the required for us values D

can be reached by the application/use only of one nonlinear cell/element in the logarithmator [5; 10], which provides the smoothness of its amplitude characteristic. Then it is possible to count that virtually always $\gamma_0 < \varepsilon$.

Furthermore, in this case error ε usually only in very the beginning and the end/lead of the amplitude characteristic noticeably differs from zero. Therefore it is possible to count that modulation indicated in practice will be noticeable only in the intervals of time, which correspond to the use of beginning and end/lead of the logarithmic characteristic (see Fig. 3a). The relative duration of these intervals of time is insignificant. Therefore in actuality the coefficient of the nonlinear distortion caused by an inaccuracy in the logarithmic characteristic, will be several (ϑ) times less than ε. Value ϑ depends on the form of the amplitude characteristic of real logarithmator. These nonlinear distortions also combination. Everything said about distortions due to narrowings is correct for them. Therefore the common/general/total equivalent coefficient of harmonics for all three forms of distortions by analogy with (12) will be recorded

$$K_{\Gamma\,\text{общ}} = \sqrt{K_\Gamma^2 + \left(\frac{\gamma_{D\,\max}}{\nu}\right)^2 + \left(\frac{\varepsilon}{\nu\vartheta}\right)^2}.$$

From expression it is evident that the specific gravity/weight of the third form of distortions is insignificant and with ε up to their several dozen percent it is possible not to consider. Consequently, requirements for the accuracy/precision of logarithmic characteristic are low.

Thus, most powerful of all three forms are the distortions, which appear as a result of the nonlinearity of logarithmic characteristic.

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Let us examine the version of logarithmic method with the application/use of compensation for the additive mixing oscillation u_3, in which the distortions of this form are strongly lowered/reduced. Let us deduct from (2) the doubled voltage "of carrier":

$$u_2 = u_1 - 2U_0 \sin(\Omega t + \Delta\psi) = -U_0 [1 - mf(t)] \sin(\Omega t + \Delta\psi). \quad (14)$$

Comparing (2) with (14), we see that the last/latter expression is also generalized AM the oscillation, which differs from the initial only in terms of the sign of the phase "of carrier". If it, it is analogous with voltage u_1, to rectify and to take the logarithm, then we will obtain

$$u_{2\,\text{лог}} = \log_N U_0 + \log_N |\sin(\Omega t + \Delta\psi)| + \log_N [1 - mf(t)]. \quad (15)$$

The last/latter term of expression (15) it is possible to expand in the following power series

$$u'_{2\,\text{лог}} = \log_N [1 - mf(t)] = -a\left[mf(t) + \frac{m^2}{2} f^2(t) + \frac{m^3}{3} f^3(t) + \frac{m^4}{4} f^4(t) + \dots\right]. \quad (16)$$

By deducting (15) from (3) and after considering (16) and (4), we **will obtain the result of the detection**

$$u_{\text{д}} = u_{1\,\text{лог}} - u_{2\,\text{лог}} = 2a\left[mf(t) + \frac{m^3}{3} f^3(t) + \frac{m^5}{5} f^5(t) + \dots\right]. \quad (17)$$

Here is straighten/rectified on resistive load and is logarithmized each of two generalized AM of oscillations u_1 and u_2, after which results they are deducted. Therefore this version can be call/named the symmetrical balance version of the logarithmic method of the generalized amplitude detection. Let us note that in (17) there are no terms with even degrees. Analogous with expression (5), it is possible to obtain **expression** for the coefficient of the harmonics of the voltage $u_{\text{д}}$, **which** in view of unwieldiness here is not given. On Fig. 4 **is represented** the graph/diagram of its dependence on m (curve 2), from which it is clear that the nonlinear distortions due to the nonlinearity of the logarithmic characteristic in symmetrical balanced version are extremely low. So, e.g., with m =

30o/o coefficient of harmonics does not exceed 1o/o. Let us focus attention on the fact that the symmetrical balance version is more complex than the unsymmetric version with the application/use of compensation, but it is very insignificant.

Thus, during the generalized amplitude detection in a logarithmic manner appear the nonlinear distortions of three forms. The distortions, which appear as a result of the nonlinearity of logarithmic characteristic, are inherent in principle in this method and in the unsymmetric version of logarithmic method they are definite.

Others two forms of distortions are the consequence of the imperfection of real logarithmator. So, the nonlinear distortions, which appear as a result of the linearity of the initial section of the amplitude characteristic of real logarithmator and which are developed in the form of the periodically being repeated narrowings of the output signal, can be decreased down to the assigned magnitude by means of an increase in the dynamic range D of the logarithmic section of the characteristic of the logarithmator.

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The desired value D being approximately two orders (in symmetrical

version) or less (in unsymmetric version), which is easily attained in practice.

The third form is the distortions, which appear as a result of an inaccuracy in the logarithmic amplitude characteristic, it is possible to consider negligible.

The total equivalent coefficient of harmonics $K_{\text{г общ}}$, that characterizes the distortions of all three forms, with unsymmetric version is approximately 10o/o, with symmetrical - in essence is determined by value D and easily can be obtained about 2-3o/o, being decreased with an increase in the "carrier" frequency Ω the initial oscillation. The latter is explained by the facts that with an increase Ω all large part of the products of the distortions of the second and third forms falls beyond the limits of passband UNCh [УНЧ - low-frequency amplifier] and therefore it is filtered out.

The distortions, which appear as a result of the nonlinearity of logarithmic characteristic, in unsymmetric version can be completely corrected by the way of the subsequent involution. Actually, if the voltage of form log U_0 [1 + m f (t)] is enforced α once, and then is fed to the input of device with exponential amplitude characteristic, then it is not difficult to show [13] that at output/yield we will obtain

$$u_{\text{вых}} = U_0' [1 + mf(t)]^{\alpha p}. \qquad (18)$$

Here U_0' is amplitude factor; $p=\ln N_э/\ln N_л$ – the relation of the natural logarithms of the foundations of the exponential and logarithmic characteristics of two inverters.

From (18) it is evident that for the absence of the nonlinearity of through amplitude characteristic it suffices to require, in order to $\alpha p = 1$. This can be accomplished with any p, making $\alpha p = 1$. It must be noted that such the addition of the logarithmic method of detection actually converts it into the logarithmic method of functional division.

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